A METHOD FOR CHANNEL EQUALIZATION, A RECEIVER, A CHANNEL EQUALIZER, AND A WIRELESS COMMUNICATION DEVICE

5   Field of the Invention

The present invention relates to a method for performing channel equalization in a receiver, in which a signal is received from a communication channel, the signal containing symbols formed of
10   binary information by phase shift keying, channel estimation is performed to estimate the properties of the communication channel, and samples are taken of the received signal at intervals. The invention also relates to a receiver comprising means for receiving a signal from a communication channel, the signal containing symbols formed of
15   binary information by phase shift keying, a channel estimator for estimating the properties of the communication channel, a channel equalizer, and means for sampling the received signal at intervals. The invention further relates to a channel equalizer comprising means for sampling a signal received from a communication channel at intervals,
20   which received signal contains symbols formed of binary information by phase shift keying, and which received signal has been subjected to channel estimation for estimating the properties of the communication channel. The invention yet relates to a wireless communication device comprising means for receiving a signal from a communication
25   channel, the signal containing symbols formed of binary information by phase shift keying, a channel estimator for estimating the properties of the communication channel, a channel equalizer, and means for sampling the received signal at intervals.

30   Background of the Invention

In radio communication, information to be transmitted is converted in a modulator to a high-frequency radio signal which is transmitted via a communication channel, i.e. a radio channel, to a receiver. In the
35   receiver, the received radio-frequency signal is demodulated, wherein the aim is to use the received high-frequency signals to form the information which substantially corresponds to the original information.

However, this communication based on radio signals is susceptible to noise which can be due to *e.g.* high-frequency signals caused by other electric devices, changes in the conditions of the communication channel, bars affecting the propagation of radio signals, such as build-
5 ings, trees, topography, *etc.* Methods have been developed to reduce the effect of such noise. In systems, in which the modulation method used is based on phase shift (M-ary Phase Shift Keying, MPSK), receivers commonly apply a channel estimator and a channel equalizer to compensate for changes caused by the communication channel in
10 the signal. The aim of a channel estimator is to determine the transfer function of the communication channel effective on the propagation of the radio signal, wherein it is possible in the channel equalizer to make a correction in the received signal on the basis of this determined transfer function. As the channel equalizer, *e.g.* the maximum
15 likelihood sequence equalizer (MLSE) and the Viterbi algorithm can be used. With such an arrangement, the effects of the communication channel can be compensated for, to at least some extent. However, such an arrangement involves *e.g.* the problem that the complexity of the system increases exponentially the higher, the more channel taps
20 are determined from the transfer function in the channel estimator. On the other hand, the more channel taps can be estimated, the better it is possible to compensate for distortions in the signal which are due to *e.g.* multipath propagation. In a formula, the complexity of the system can be depicted by M^H, in which H is the number of estimated channel
25 taps and M is the number of different phase-shift alternatives to be used in the modulation.

As the radio signal is reflected from obstacles, such as buildings, vegetation, topography, *etc.*, it causes so-called multipath propagation,
30 wherein the same signal comes to the receiver along several different paths. Since these different signal paths can have different lengths and they are longer than the distance travelled by the signal directly from the transmitter to the receiver, the signals propagating along different paths are received by the receiver at different times. Furthermore, such
35 a multipath-propagated signal can be stronger than the directly propagated signal, wherein signals which have come to the receiver along different paths cannot be differentiated from each other solely on

the basis of the signal strength. Particularly in said phase modulation, multipath propagation has a significant effect on the performance of the receiver. In principle, for example four-tap estimation can be used to compensate for four multipath-propagated signals which come to the receiver at different times. However, receivers of prior art are thus very complex compared to a situation in which for example one-tap estimation is used.

The following is a brief discussion of the transmission of binary information by using phase shift keying. Let us fix the time index t to $t = \tau + t\Delta\tau$ in which $\Delta\tau$ is the symbol transmission interval. The symbol $S(B_t)$ corresponds to three bits $B_t = [b_{t,1}, b_{t,2}, b_{t,3}]$ at time t ($b_{t,k} \in \{0,1\}$), when 8PSK modulation is applied. Thus, the symbol generator can be presented by the formula

$$
\begin{aligned}
S(B) &= a^2(1-b_1)b_2 b_3 + a^4(1-b_1)b_2(1-b_3) \\
&+ a^6(1-b_1)(1-b_2)(1-b_3) + a^8(1-b_1)(1-b_2)b_3 \\
&+ a^{10}b_1(1-b_2)b_3 + a^{12}b_1(1-b_2)(1-b_3) \\
&+ a^{14}b_1 b_2(1-b_3) + a^{16}b_1 b_2 b_3
\end{aligned}
\tag{1}
$$

where $a = e^{i2\pi/16}$.

The symbols are transmitted to the communication channel. In the receiver, the signal to be received at time t is

$$
r_t = \sum_{s=0}^{H-1} \tilde{h}_s S(B_{t-s}) + n_t
\tag{2}
$$

in which $\tilde{h}_s$ indicates H channel taps and n_t indicates channel noise which is assumed to be Gaussian noise.

An optimal channel equalizer would be one that maximizes the likelihood function. If it is assumed that the noise mixed with the signal is Gaussian noise and the probability of all the symbols S is equal, then the maximization of the likelihood functions corresponds to the minimization of the function

$$f(B) = \sum_{t=0}^{T} \left\| r_t - \sum_{s=0}^{H-1} h_s S(B_{t-s}) \right\|^2 \qquad (3)$$

in which $S(B)$ is the symbol corresponding to bits B, h_s are the
estimated channel coefficients, and r is the received signal. Conse-
quently, the problem is to find the bits B after receiving r and estimat-
ing h. In the above formula, $T+1$ is the number of transmitted symbols,
H indicates the number of channel taps, and s is the tap index. To
simplify the notations, time index t indicates the value $u+t\Delta u$ in which u
is time and Δu is the sampling interval. The function f can be minimized
by using the Viterbi algorithm. However, the problem is that the
complexity of the Viterbi algorithm increases exponentially as the
function of the number of channel taps H, as already mentioned above.
Another method is to first use a prefilter to reduce the number of
channel taps to for example two and, after this, to apply the Viterbi
algorithm to maximize the likelihood function. However, this method
has the drawback that the result depends *e.g.* on the quality of the
prefilter and the type of the channel. One method applying the Viterbi
algorithm is based on the use of a delayed decision-feedback
sequence estimator (DDFSE). In this method, the Viterbi algorithm is
applied in the feedback filter. The performance of such a sequence
estimator is affected *e.g.* the number of states used in Trellis coding.
If the number of states is sufficiently large, the performance is close to
the optimum, but the estimator is thus very complex. Such an estimator
operates the better, the smaller the phase shift in the channel, wherein
the estimator requires a filter whereby this phase shift can be reduced.

Summary of the Invention

It is an aim of the present invention to provide an improved method for
channel equalization, as well as a channel equalizer. To apply the
method of the invention, an iterative algorithm has been developed to
minimize the function f. More precisely, the method according to the
present invention is primarily characterized in that in the method, a
determined number of samples are examined, a decision step is taken,
in which, to find out the transmitted symbols, the bit decisions are

computed on the basis of said defined quantity of samples, and after each decision step it is examined whether said decision step is to be iterated, wherein upon iteration of said decision step, at least some of the bit decisions of the previous decision step are used in addition to the samples under examination, in the computation of the bit decision. The receiver according to the present invention is primarily characterized in that the channel equalizer comprises means for examining a number of samples defined at the time, decision means for computing bit decisions on the basis of said defined number of samples to find out the transmitted symbols, and examining means for estimating the need for iterating the computation of the bit decisions, wherein upon iterating said computation of bit decisions, at least some of the bit decisions of the previous decision step are arranged to be used in addition to the samples under examination at the time. The channel equalizer according to the present invention is primarily characterized in that the channel equalizer comprises means for examining a number of samples defined at a time, decision means for computing bit decisions on the basis of said defined number of samples to find out the transmitted symbols, and examining means for estimating the need for iterating the computation of the bit decisions, wherein upon iterating said computation of bit decisions, at least some of the bit decisions of the previous decision step are arranged to be used in addition to the samples under examination at the time. Further, the wireless communication device according to the present invention is primarily characterized in that the channel equalizer comprises means for examining a number of samples defined at the time, decision means for computing bit decisions on the basis of said defined number of samples to find out the transmitted symbols, and examining means for estimating the need for iterating the computation of the bit decisions, wherein upon iterating said computation of bit decisions, at least some of the bit decisions of the previous decision step are arranged to be used in addition to the samples under examination at the time.

The present invention shows remarkable advantages when compared to solutions of prior art. Applying the method of the invention, the structure of the channel equalizer is simpler than with prior art

solutions, as the number of channel taps is increased. The channel equalizer according to the invention can also be implemented as an integrated circuit, wherein the size of the device applying the channel equalizer can be reduced and its manufacture is less expensive than

5 the manufacture of electronic devices applying channel equalizers of prior art. The simpler implementation also provides that the power consumption can be reduced, which is particularly advantageous in portable electronic devices.

10 Brief Description of the Drawings

In the following, the invention will be described in more detail with reference to the appended drawings, in which

15 Fig. 1   shows an electronic device, in connection with which the channel equalizer according to the invention can be applied,

Fig. 2a   shows a channel equalizer according to a preferred embodiment of the invention in a reduced block chart,

20 Fig. 2b   shows a computing block for use in the channel equalizer of Fig. 2a, in a reduced block chart, and

Fig. 3   illustrates comparison of simulation results obtained with
25   different types of channel equalizers.

Detailed Description of the Invention

In the following description of the method according to an advanta-
30 geous embodiment of the invention, a wireless communication device with a receiver 2 and a transmitter 3 for radio-frequency communication will be used as an example of an electronic device 1. As the modulation method, M-ary phase shift keying M PSK will be used, such as 8 PSK (M = 8). In such phase shift keying, eight different phase shift
35 values are available for symbol formation, wherein one symbol can be used to transmit the information of three bits (2^3). Generally in phase shift keying, the letter M indicates the involution of two 2^N, in which N is

greater than 0. Such an 8 PSK modulated signal has been proposed for example for the GSM-EDGE system, but it is obvious that the invention can also be applied in connection with other phase shift modulation methods and other communication systems applying phase
5   shift modulation.

The receiver 2 comprises a high-frequency part 4, in which the received high-frequency signal is converted either to an intermediate-frequency signal or directly to the baseband. The signal formed in the
10   high-frequency part 4 is converted to digital format by sampling in an analog-to-digital converter 5. These samples are stored in a memory 10 for further processing steps. On the basis of these samples, the receiver 2 is synchronized with the received signal in a synchronization block 6. Here, so-called training sequences can be utilized, which are
15   transmitted with payload information to the receiver 2. Such a training sequence and its location in relation to the payload information is also known in the receiver 2, wherein the receiver can search for these training sequences in the received signal *e.g.* by a correlation method. The synchronization block 6 can also be used to control the timing of
20   the analog-to-digital converter 5. An automatic amplification control block 8 can be used to control the amplification of the high-frequency part 4, if necessary, wherein variations in signal strength can be compensated for, to at least some extent. The received, digitized signal is transmitted to a channel equalizer 7 in which the received signal is
25   subjected to equalizing operations to compensate, as well as possible, for changes in signal propagation due to the radio channel. The channel equalizer 7 also comprises a channel estimator 17 (Fig. 2a) for estimating the properties of the communication channel, *i.e.* for finding out the transfer function of the communication channel. Subsequently,
30   the channel-coded signal is directed to a decoder 9 for decoding. The function of the decoder 9 is to find out from the received signal the original information transmitted in the signal, after which the received information is led to further processing steps in a way known as such.

35   The following is a mathematical description of the computing algorithm to be used in the method according to the invention. For clarity, it is assumed in the description that 8 PSK modulation ($M = 3$) is involved,

wherein the symbol value is determined by three bits, but the principles to be presented below can also be applied in other M PSK encodings. The purpose of the computing algorithm is to minimize the function $f(B)$ presented in formula (3) in an iterative manner. To facilitate the description, two terms are added in the cost function (3), resulting in the cost function

$$f(B) = \frac{1}{2} \sum_{t=0}^{T} \left(\left\| r_t - \sum_{s=0}^{H-1} h_s S(B_{t-s}) \right\|^2 - \sum_{s=0}^{H-1} \overline{h}_s h_s \overline{S}(B_{t-s}) S(B_{t-s}) + \sum_{k=1}^{M} \left(b_{l,k} - \frac{1}{2} \right)^2 \right) \quad (4)$$

in which the reference $\overline{x}$ indicates the complex conjugate of x. Since the terms b are binary, they have a constant value. Thus, the minimum of the function (4) corresponds to the minimum of the function (3). Taking the derivative of $f(B)$ according to formula (4) with respect to $b_{l,k}$ ($l = t - s$), we get that

$$\frac{\Delta f(B)}{\Delta b_{l,k}} = b_{l,k} - \frac{1}{2} + \sum_{t=l}^{l+H-1} \left(-re\left\{ \overline{r}_t h_{t-l} \frac{\Delta S(B_l)}{\Delta b_{l,k}} \right\} + re\left\{ \overline{h}_{t-l} \frac{\Delta \overline{S}(B_l)}{\Delta b_{l,k}} \sum_{q=0,t-q \neq l}^{H-1} h_q S(B_{t-q}) \right\} \right) \quad (5)$$

in which the reference $re\{x\}$ indicates the real part of x.

Now, using the gradient descent and requiring that b is either 0 or 1, we get that

$$b_{l,k}(j+1) = g_h\left(b_{l,k}(j) - \alpha \frac{\Delta f(B_l)}{\Delta b_{l,k}} \right) \quad (6)$$

where j refers to iteration number and $g_h(x) = 1$, if $x > 0.5$, else 0.

Setting $\alpha = 1$, we get the update rule for the bit $b_{l,k}$, ($k = 1, 2, 3$; l is an index):

$$b_{l,k}(j+1) = f_h\left(\sum_{t=l}^{l+H-1} re\left\{ \overline{r}_t h_{t-l} \frac{\Delta S(B_l)}{\Delta b_{l,k}} \right\} - re\left\{ \overline{h}_{t-l} \frac{\Delta \overline{S}(B_l)}{\Delta b_{l,k}} \sum_{q=0,t-q \neq l}^{H-1} h_q S(B_{t-q}) \right\} \right) \quad (7)$$

where $B_l = \lfloor b_{l,1}, b_{l,2}, b_{l,3} \rfloor$ are three bits at time $l = u + l\Delta u$, $S(B_l)$ is the corresponding symbol, $S(B_l)$ is the derivative with respect to k bits, h indicates the communication channel, of which H channel taps are estimated (H-path channel), and $f_h(x)$ is a hard limit function which receives the value 1, if $x > 0$, else 0. In the general case, $B_l = \lfloor b_{l,1}, b_{l,2}, ..., b_{l,M} \rfloor$. The symbols S are complex numbers, and $re\{\}$ indicates taking the real part. To compute the iteration step to be taken by the update rule according to this function, it is possible to select how long a signal is to be examined. For example, a period of 5H can be examined, i.e. $l = 0,...,5H-1$.

The description S is defined, as presented in formula (1), in such a way that

$$\begin{aligned} S(B_l) &= a^2\left(1-b_{l,1}\right)b_{l,2}b_{l,3} + a^4\left(1-b_{l,1}\right)b_{l,2}\left(1-b_{l,3}\right) \\ &+ a^6\left(1-b_{l,1}\right)\left(1-b_{l,2}\right)\left(1-b_{l,3}\right) + a^8\left(1-b_{l,1}\right)\left(1-b_{l,2}\right)b_{l,3} \\ &+ a^{10}b_{l,1}\left(1-b_{l,2}\right)b_{l,3} + a^{12}b_{l,1}\left(1-b_{l,2}\right)\left(1-b_{l,3}\right) \\ &+ a^{14}b_{l,1}b_{l,2}\left(1-b_{l,3}\right) + a^{16}b_{l,1}b_{l,2}b_{l,3} \end{aligned} \qquad (8)$$

where $a = \exp(i2\pi l/16)$. The bit values are initialized for example randomly. Because binary optimization is involved, the method may be stuck in a local minimum. However, this can be prevented by adding noise whose variance is reduced during convergence, to the iteration. Furthermore, several initializations can be used.

The structure of a channel equalizer 7 complying with a preferred embodiment of the invention is shown in the appended Fig. 2a. To start the operation, the states $b_{l,k}$ are set for example randomly. The sampled signal r which is delayed in delay blocks 15a–15d of a delay line, is led to the channel equalizer 7. Furthermore, channel coefficients are estimated from the signal in the channel estimator 17 which forms a vector h. After the required number (in this example 5H) of samples has been led to the channel equalizer 7, the iteration step is started. Thus, the values of the vector h as well as the sample values of the sampled signal r, delayed in the delay blocks 15a–15d, are input in computing units 16a–16e, one of which is shown in more detail in

Fig. 2b. Further, the bit decisions B of the neighbouring computing units, whose values at the first iteration time are the values determined at the initialization stage, are input in the computing unit.

5 The computing units 16a–16e contain a number corresponding to the term M of iteration blocks, in this example three iteration blocks 18a, 18b, 18c, each of which implements the iteration step according to the above formula (7), except for the hard limit function f_h, that is

$$10 \quad \tilde{b}_{l,k}(j+1) = \sum_{t=l}^{l+H-1} re\left\{\overline{r}_t h_{t-l} \frac{\Delta S(B_l)}{\Delta b_{l,k}}\right\} - re\left\{\overline{h}_{t-l} \frac{\Delta \overline{S}(B_l)}{\Delta b_{l,k}} \sum_{q=0,t-q\neq l}^{H-1} h_q S(B_{t-q})\right\} + AWGN \quad (9)$$

where $k = 1, 2, 3$ and $AWGN$ refers to Gaussian noise. The hard limit function is implemented by means of hard limit blocks 19a, 19b, 19c. In the references of Fig. 2b, the character ~ on top of a symbol indicates

15 the decision value before performing the hard limit function. The iteration blocks 18a, 18b, 18c are coupled to each other in such a way that their output is led to the other iteration blocks 18a, 18b, 18c in the same computing unit 16a–16e preferably through the hard limit blocks 19a, 19b, 19c, but not to the input of the same iteration block; that is,

20 there is no feedback in the iteration blocks 18a–18c. The block n forms random noise which is input in the iteration blocks 18a, 18b, 18c of the computing unit, if necessary. Preferably, the variance of the added noise should reduce during the step of stabilization. Furthermore, the output of all the three limiting blocks 19a, 19b, 19c in the computing

25 block is preferably input in at least one, preferably neighbouring computing unit 16a–16e. The number of computing units 16a–16e in which the output is input, depends on the number of channel taps. After the computing blocks 16a–16e have computed the bit value of one iteration step, the hard limit function f_h is computed in the hard limit

30 block 19a, 19b, 19c which makes the bit decisions B of the computing blocks at the respective iteration round. After this, it is examined, for example in the control block 11, whether the state of the channel equalizer 7 is stabilized, or if there is still a need to iterate the computation of the bit decisions. If it is noticed that the state is not yet

35 sufficiently stable, a new round of iteration computing is performed, in

which the bit decisions formed in the previous iteration round are now used as the bit decisions B. Stabilization can be examined for example by comparing the bit decisions of the previous iteration round to the bit decisions of the latest iteration round, wherein if the bit decisions have remained the same, it can be assumed that the channel equalizer 7 is stabilized.

After the channel equalizer has been stabilized, a channel-equalized binary signal can be obtained from the hard limit blocks of the channel equalizer, for further processing. Thus, the output of each hard limit block contains the bit decision B_i which is a vector containing the channel-equalized binary values corresponding to the symbol r_i.

From the computing units 16a–16e it is also possible to take so-called soft decisions $\tilde{B}_0$ by using an activation function f_a which can be for example a sigmoid function. Soft decisions are advantageous in decoding of bits.

The channel equalizer 7 shown in Figs. 2a and 2b can be primarily implemented by programming *e.g.* in a digital signal processor (DSP) which can be *e.g.* a part of the control block 11. Furthermore, a memory is needed in the operation of the channel equalizer 7, for storing the bit decisions made at the iteration rounds. The channel equalizer 7 can also be implemented as an integrated circuit.

Further, Fig. 3 shows simulation results comparing the method according to an advantageous embodiment of the invention with methods of prior art. The method according to an advantageous embodiment of the invention has been tested with a complex 4-path channel. The channel coefficient h is estimated from a training sequence of 100 bits. Also, parametres of a delayed decision-feedback sequence estimator (DDFSE) according to prior art are estimated from the corresponding sequence. Noise has been added during the iteration. The result shows that in this test, the new method (curve A) is clearly better than the reference method (curve B). The number of iterations was 50. The dotted line (curve C) shows the result obtained

with the method according to the invention by initializing the bit values b with the real values, by eliminating the added noise and letting the equalizer stabilize. This is very close to the performance of the Viterbi algorithm.

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It is obvious that the present invention is not limited solely to the above-presented embodiments, but it can be modified within the scope of the appended claims.